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Chapter 21

SUPER-HIGH FREQUENCY MODELS AND BEHAVIOUR OF IMs

21.1 INTRODUCTION

Voltage strikes and restrikes produced during switching operations of induction motors fed from standard power grid may cause severe dielectric stresses on the stator IM windings, leading, eventually, to failure.

In industrial installations high dielectric stresses may occur during second and third pole circuit breaker closure. The second and the third pole closure in electromagnetic power circuit breakers has been shown to occur within 0 to 700 μs [1].

For such situations, electrical machine modelling in the frequency range of a few KHz is required. Steep fronted waves with magnitudes up to 5 p.u. may occur at the machine terminals under certain circuit breaker operating conditions.

On the other hand, PWM voltage source inverters produce steep voltage pulses which are applied repeatedly to induction motor terminals in modern electric drives.

In IGBT inverters, the voltage switching rise times of 0.05–2 μs , in presence of long cables, have been shown to produce strong winding insulation stresses and premature motor bearing failures.

With short rise time IGBTs and power cables longer than a critical l_c , repetitive voltage pulse reflection may occur at motor terminals.

The reflection process depends on the parameters of the feeding cable between motor and inverter, the IGBTs voltage pulse time t_r , and the motor parameters.

The peak line to line terminal overvoltage (V_{pk}) at the receiving end of an initially uncharged transmission line (power cable) subjected to a single PWM pulse with rise time t_r [2] is

$$\begin{aligned} (V_{pk})_{l \geq l_c} &= (1 + \Gamma_m) V_{dc} \\ \Gamma_m &= \frac{Z_m - Z_0}{Z_m + Z_0} \end{aligned} \quad (21.1)$$

where the critical cable length l_c corresponds to the situation when the reflected wave is fully developed; V_{dc} is the dc link voltage in the voltage source inverter and Γ_m is the reflection coefficient ($0 < \Gamma_m < 1$).

Z_0 is the power cable and Z_m – the induction motor surge impedance. The distributed nature of a long cable L–C parameters favor voltage pulse reflection,

besides inverter short rising time. Full reflection occurs along the power cable if the voltage pulses take longer than one-third the rising time to travel from converter to motor at speed $u^* \approx 150\text{--}200\text{ m}/\mu\text{s}$.

The voltage is then doubled and critical length is reached. [2] (Figure 21.1)

The receiving (motor) end may experience $3V_{dc}$ for cable lengths greater than l_c when the transmission line (power cable) has initial trapped charges due to multiple PWM voltage pulses.

Inverter rise times of $0.1\text{--}0.2\text{ }\mu\text{s}$ lead to equivalent frequency in the MHz range.

Consequently, investigating super-high frequency modelling of IMs means frequency from kHz to 10MHz or so.

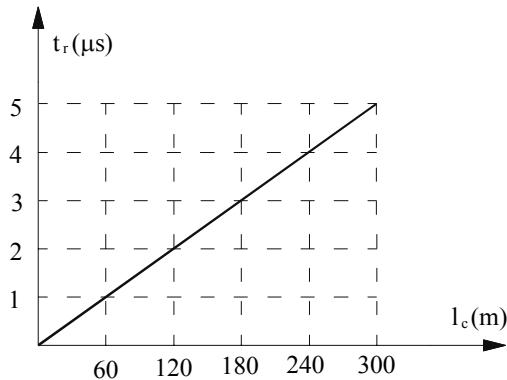


Figure 21.1 Critical power cable length l_c

The effects of such fast voltage pulses on the machine windings include:

- Non-uniform voltage distribution along the windings, with the first 1–2 coils (connected to the terminals) experiencing most of the voltage drop. Especially with random coils, the inter-turn voltage between the first and, say, the last turn, which may be located nearby, may become high enough to produce premature insulation aging.
- The common mode voltage PWM inverter pulses, on the other hand, produce parasitic capacitive currents between stator windings and motor frame and, in parallel, through airgap parasitic capacitance and through bearings, to motor frame. The common mode circuit is closed through the cable capacitances to ground.
- Common mode current may unwarrantably trip the null protection of the motor and damage the bearing by lubricant electrostatic breakdown.

The super-high frequency or surge impedance of the IM may be approached either globally when it is to be identified through direct tests or as a complex

distributed parameter (capacitor, inductance, resistance) system, when identification from tests at the motor terminals is in fact not possible.

In such a case, either special tests are performed on a motor with added measurement points inside its electric (magnetic) circuit, or analytical or FEM methods are used to calculate the distributed parameters of the IM.

IM modelling for surge voltages may then be used to conceive methods to attenuate reflected waves and change their distribution within the motor so as to reduce insulation stress and bearing failures.

We will start with global (lumped) equivalent circuits and their estimation and continue with distributed parameter equivalent circuits.

21.2 THREE HIGH FREQUENCY OPERATION IMPEDANCES

When PWM inverter-fed, the IM terminals experience three pulse voltage components.

- Line to line voltages (example: phase A in series with phase B and C in parallel): V_{ab} , V_{bc} , V_{ca}
- Line (phase) to neutral voltages: V_{an} , V_{bn} , V_{cn}
- Common mode voltage V_{oin} (Figure 21.2)

$$V_{oin} = \frac{V_a + V_b + V_c}{3} \quad (21.2)$$

Assume that the zero sequence impedance of IM is Z_0 .

The zero sequence voltage and current are then

$$V_0 = \frac{(V_a - V_n) + (V_b - V_n) + (V_c - V_n)}{3} = \frac{V_a + V_b + V_c}{3} - V_n \quad (21.3)$$

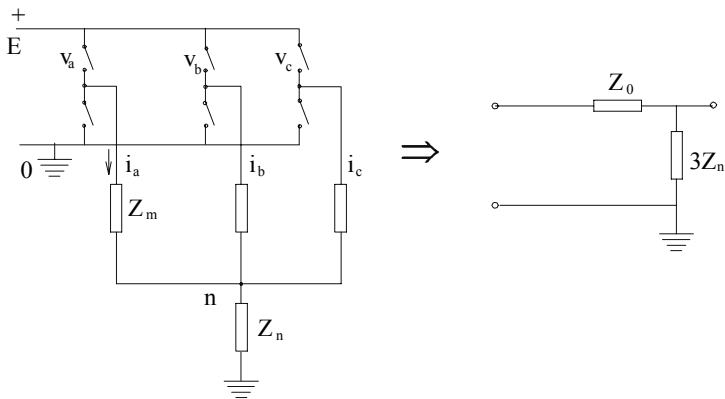


Figure 21.2 PWM inverter with zero sequence impedance of the load (insulated neutral point)

Also, by definition,

$$\begin{aligned} V_n &= Z_n I_n \\ V_0 &= Z_0 I_0 \\ I_0 &= \frac{I_a + I_b + I_c}{3} = \frac{I_n}{3} \end{aligned} \quad (21.4)$$

Eliminating V_n and V_0 from (21.3) with (21.4) yields

$$I_n = \frac{3}{Z_0 + 3Z_n} \cdot \frac{(V_a + V_b + V_c)}{3} \quad (21.5)$$

and again from (21.3),

$$\begin{aligned} V_{oin} &= \frac{V_a + V_b + V_c}{3} = Z_0 I_0 + V_n \\ V_n &= 3Z_n I_0 = Z_n I_n \end{aligned} \quad (21.6)$$

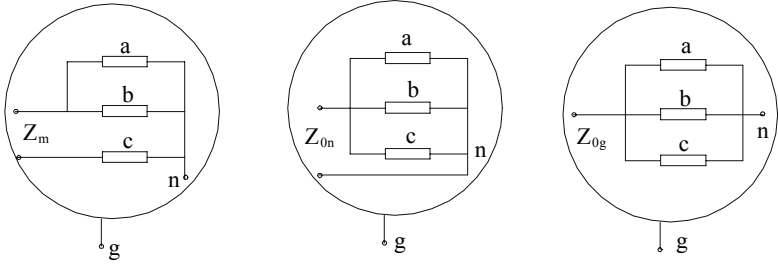


Figure 21.3 Line (Z_m) neutral (Z_{on}) and ground (common mode Z_{og})

This is how the equivalent lumped impedance (Figure 21.2) for the common voltage evolved.

It should be noted that for the differential voltage mode, the currents flow between phases and thus no interference between the differential and common modes occurs.

To measure the lumped IM parameters for the three modes, terminal connections as in Figure 21.3 are made.

Impedances in Figure 21.3, dubbed here as differential (Z_m), zero sequence – neutral – (Z_{on}), and common mode (Z_{og}), may be measured directly by applying single phase ac voltage of various frequencies. Both the amplitude and the phase angle are of interest.

21.3 THE DIFFERENTIAL IMPEDANCE

Typical frequency responses for the differential impedance: Z_m are shown in Figure 21.4. [3]

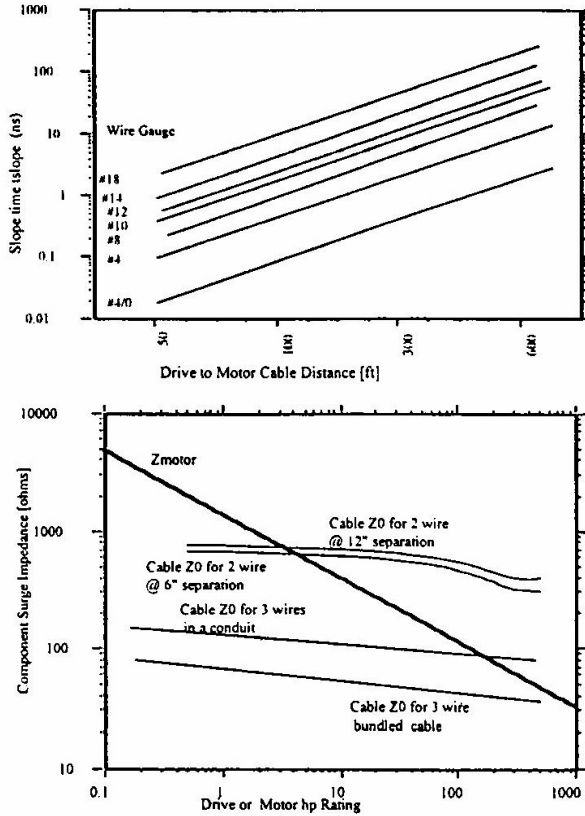
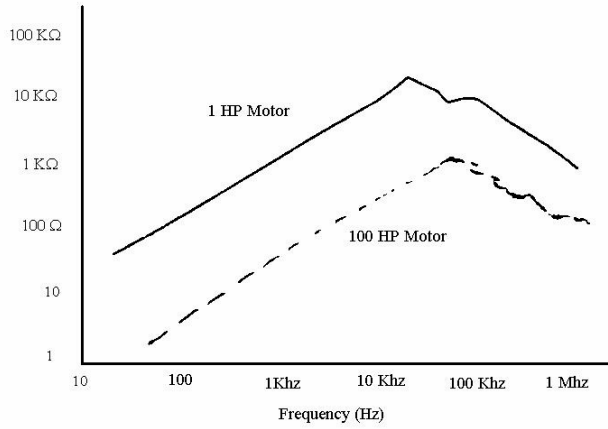


Figure 21.4 Differential mode impedance (Z_m)

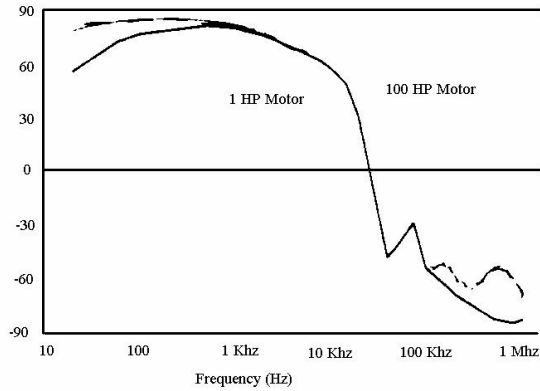
Z_m has been determined by measuring the line voltage V_{ac} during the PWM sequence with phase a,b together and c switched from the + d.c. bus to the - d.c. bus. The value Δe is the transient peak voltage V_{ac} above d.c. bus magnitude during t_{rise} and ΔI is phase c peak transient phase current.

$$Z_m = \frac{\Delta e}{\Delta I} \quad (21.7)$$

As seen in Figure 21.4, Z_m decreases with IM power. Also it has been found out that Z_m varies from manufacturer to manufacturer, for given power, as much as 5/1 times. When using an RLC analyser (with phases a and b in parallel, connected in series with phase c) the same impedance has been measured by a frequency response test (Figure 21.5). [3]



a.)



b.)

Figure 21.5 Differential impedance Z_m versus frequency
a.) amplitude b.) phase angle

The phase angle of the differential impedance Z_m approaches a positive maximum between 2 kHz and 3 kHz (Figure 21.4 b). This is due to lamination skin effect which reduces the iron core ac inductance.

At critical core frequency f_{iron} , the field penetration depth equals the lamination thickness d_{lam} .

$$d_{lam} = \sqrt{\frac{1}{\pi f_{iron} \sigma_{iron} \cdot \mu_{rel} \cdot \mu_0}} \quad (21.8)$$

The relative iron permeability μ_{rel} is essentially determined by the fundamental magnetization current in the IM. Above f_{iron} eddy current shielding becomes important and the iron core inductance starts to decrease until it approaches the wire self-inductance and stator aircore leakage inductance at the resonance frequency $f_r = 25$ kHz (for the 1 HP motor) and 55 kHz for the 100 HP motor.

Beyond f_r , turn to turn and turn to ground capacitances of wire perimeter as well as coil to coil and phase to phase capacitances prevail such that the phase angle approaches now -90° degrees (pure capacitance) around 1 MHz.

So, as expected, with increasing frequency the motor differential impedance switches character from inductive to capacitive. Z_m is important in the computation of reflected wave voltage at motor terminals with the motor fed from a PWM inverter through a power (feeding) cable. Many simplified lumped equivalent circuits have been tried to model the experimentally obtained wide-band frequency response [4,5].

As very high frequency phenomena are confined to the stator slots, due to the screening effect of rotor currents – and to stator end connection conductors, a rather simple line to line circuit motor model may be adopted to predict the line motor voltage surge currents.

The model in [3] is basically a resonant circuit to handle the wide range of frequencies involved (Figure 21.6)

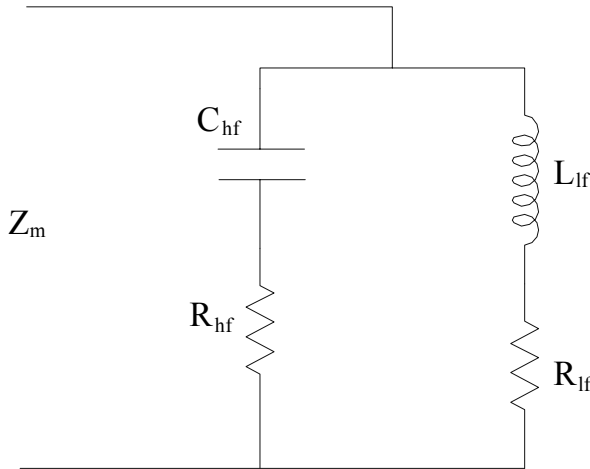


Figure 21.6 Differential mode IM equivalent circuit

C_{hf} and R_{hf} determine the model at high frequencies (above 10kHz in general) while L_{lf} , R_{lf} are responsible for lower frequency modelling. The identification of the model in Figure 21.6 from frequency response may be done through optimization (regression) methods.

Typical values of C_{hf} , R_{hf} , L_{lf} , R_{lf} are given in Table 21.1 after [3].

Table 21.1 Differential mode IM model parameters

	1 kW	10 kW	100 kW
C_{hf}	250pF	800pF	8.5nF
R_{hf}	18Ω	1.3Ω	0.13Ω
R_{lf}	150Ω	300Ω	75Ω
L_{lf}	190mH	80mH	3.15μH

Table 21.1 suggests a rather linear increase of C_{hf} with power and a rather linear decrease of R_{hf} , R_{lf} and L_{lf} with power.

The high frequency resistance R_{hf} is much smaller than the low frequency resistance R_{lf} .

21.4 NEUTRAL AND COMMON MODE IMPEDANCE MODELS

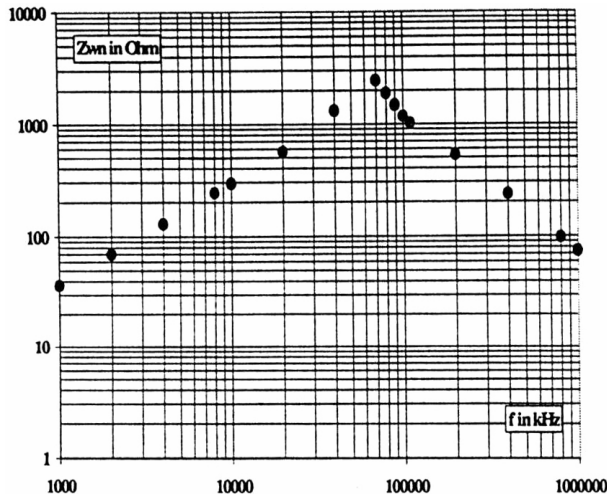
Again we start our inquiries from some frequency response tests for the connections in Figure 21.3 b (for Z_{on}) and in Figure 21.3 c (for Z_{og}) [6]. Sample results are shown in Figure 21.7 a,b.

Many lumped parameter circuits to fit results such as those in Figure 21.7 may be tried.

Such a simplified phase circuit is shown in Figure 21.8 [6].

The impedances Z_{on} and Z_{og} with the three phases in parallel from Figure 21.8 are

$$Z_{on} = \frac{pL_d}{3 \left[1 + p \frac{L_d}{R_e} + p^2 L_d \frac{C_g}{2} \right]} \quad (21.9)$$



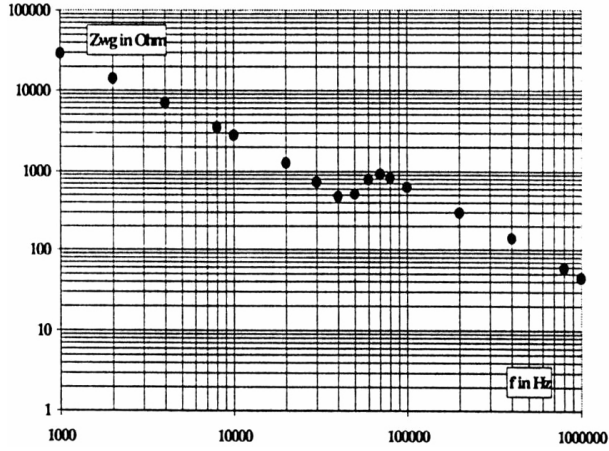


Figure 21.7. Frequency dependence of Z_{on} and Z_{og}

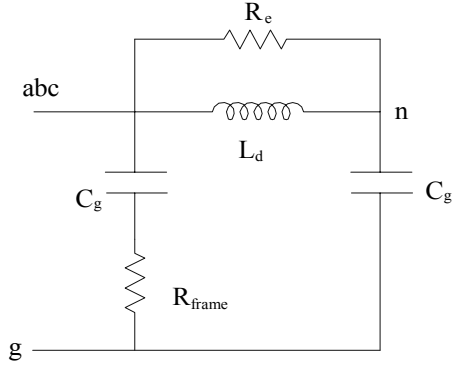


Figure 21.8 Simplified high frequency phase circuit

$$Z_{og} = \frac{1 + p \frac{L_d}{R_e} + p^2 L_d C_g}{6pC_g \left[1 + p \frac{L_d}{R_e} + p^2 L_d \frac{C_g}{2} \right]} \quad (21.10)$$

The poles and zeros are

$$f_p(Z_{on}, Z_{og}) = \frac{1}{2\pi} \sqrt{\frac{2}{L_d C_g}} \quad (21.11)$$

$$f_z(Z_{og}) = f_p(Z_{on}, Z_{og}) / \sqrt{2} \quad (21.12)$$

At low frequency (1 kHz – 10 kHz), Z_{og} is almost purely capacitive.

$$C_g = \frac{1}{6 \times 2\pi f \times (Z_{og})_f} \quad (21.13)$$

With $f = 1$ kHz, and Z_{og} (1 kHz) from [Figure 21.7 b](#), the capacitance C_g may be determined.

With the pole frequency f_p (Z_{on} , Z_{og}), corresponding to peak Z_{on} value, the resistance R_e is

$$R_e = 3(Z_{on})_{f_p} \quad (21.14)$$

Finally, L_d is obtained from (21.11) with f_p known from [Figure 21.7](#) and C_g calculated from (21.13).

This is a high frequency inductance.

It is thus expected that at low frequency (1 – 10 kHz) fitting of Z_{on} and Z_{og} with the model will not be so good.

Typical values for a 1.1 kW 4 pole, 50 Hz, 220/380 V motor are $C_g = 0.25$ nF, $L_d(h_f) = 28$ mH, $R_e = 17.5$ k Ω .

For 55 kW $C_g = 2.17$ nF, $L_d(h_f) = 0.186$ mH, $R_e = 0.295$ k Ω .

On a logarithmic scale, the variation of C_g and $L_d(h_f)$ with motor power in kW may be approximated [6] to

$$C_g = 0.009 + 0.53 \ln(P_n(\text{kW})) , \text{ [nF]} \quad (21.15)$$

$$\ln(L_d(h_f)) = 2.36 - 0.1P_n(\text{kW}) , \text{ [mH]} \quad (21.16)$$

for 220/380 V, 50 Hz, 1 – 55 kW IMs.

To improve the low frequency between 1 kHz and 10 kHz fitting between the equivalent circuit and the measured frequency response, an additional low frequency R_e , L_e branch may be added in parallel to L_d in [Figure 21.8](#).

Alternatively, the dq model may be placed in parallel and thus obtain a general equivalent circuit acceptable for digital simulations at any frequency.

The addition of an eddy current resistance representing the motor frame R_{frame} ([Figure 21.8](#)) may also improve the equivalent circuit precision.

Test results with triangular voltage pulses and with a PWM converter and short cable have proven that the rather simple high frequency phase equivalent circuit in [Figure 21.8](#) is reliable.

More elaborated equivalent circuits for both differential and common voltage modes may be adopted for better precision. [5] For their identification however, regression methods have to be used.

The lumped equivalent circuits for high frequency presented here are to be identified from frequency response tests. Their configuration retains a large degree of approximation. They serve only to assess the impact of differential and common mode voltage surges at the induction motor terminals.

The voltage surge paths and their distribution within the IM will be addressed below.

21.5 THE SUPER-HIGH FREQUENCY DISTRIBUTED EQUIVALENT CIRCUIT

When the power grid connected, IMs undergo surge-connection or atmospherical surge voltage pulses. When PWM voltage surge-fed, IMs experience trains of steep front voltage pulses. Their distribution, in the first few microseconds, along the winding coils, is not uniform.

Higher voltage stresses in the first 1–2 terminal coils, and their first turns occur.

This non uniform initial voltage distribution is due to the presence of stray capacitors between turns (coils) and the stator frame.

The complete distributed circuit parameters should contain individual turn to turn and turn to ground capacitances, self – turn, turn-to-turn and coil-to-coil inductances and self-turn, eddy current resistances.

Some of these parameters may be measured through frequency response tests only if the machine is tapped adequately for the purpose. This operation may be practical for a special prototype to check design computed values of such distributed parameters.

The computation process is extremely complex even impossible in terms of turn-to-turn parameters in random wound coil windings.

Even via 3D–FEM, the complete set of distributed circuit parameters valid from 1 kHz to 1 MHz, is not yet feasible.

However, at high frequencies (in the MHz range), corresponding to switching times in the order of tens of a microsecond, the magnetic core acts as a flux screen and thus most of the magnetic flux will be contained in air as leakage flux.

The high frequency eddy currents induced in the rotor core will confine the flux within the stator. In fact the magnetic flux will be non-zero in the stator slot and in the stator coil end connections zone.

Let us suppose that the end connections resistances, inductances, and capacitances between various turns and to the frame can be calculated separately. The skin effect is less important in this zone and the capacitances between the end turns and the frame may be neglected, as their distance to the frame is notably larger in comparison with conductors in slots which are much closer to the slot walls.

So, in fact, the FEM may be applied within a single stator slot, conductor by conductor ([Figure 21.9](#)).

The magnetic and electrostatic field is zero outside the stator slot perimeter Γ and non zero inside it.

The turn in the middle of the slot has a lower turn-to-ground capacitance than turns situated closer to the slot wall.

The conductors around a conductor in the middle of the slot act as an eddy current screen between turn and ground (slot wall).

In a random wound coil, positions of the first and of the last turns (n_c) are not known and thus may differ in different slots.

Computation of the inductance and resistance slot matrixes is performed separately within an eddy current FEM package. [7]

The first turn current is set to 1A at a given high frequency while the current is zero in all other conductors. The mutual inductances between the active turn and the others is also calculated.

The computation process is repeated for each of the slot conductors as active.

With $I = 1A$ (peak value), the conductor self inductance L and resistance R are

$$L = \frac{2W_{\text{mag}}}{\left(\frac{I_{\text{peak}}}{\sqrt{2}}\right)^2}; R = \frac{P}{\left(\frac{I_{\text{peak}}}{\sqrt{2}}\right)^2} \quad (21.17)$$

The output of the eddy current FE analysis per slot is an $n_c \times n_c$ impedance matrix.

The magnetic flux lines for $n_c = 55$ turns/coil (slot), $f \approx \frac{1}{t_{\text{rise}}}$, from 1 to 10 MHz, with one central active conductor ($I_{\text{peak}} = 1A$) (Figure 21.9, [7]) shows that the flux lines are indeed contained within the slot volume.

The eddy current field solver calculates A and Φ in the field equation

$$\nabla_x \frac{1}{\mu_r} (\nabla \times A) = (\sigma + j\omega\epsilon_r) (-j\omega A - \Delta\Phi) \quad (21.18)$$

with:

$A(x,y)$ – the magnetic potential (Wb/m)

$\Phi(x,y)$ – the electric scalar potential (V)

μ_r – the magnetic permeability

ω – the angular frequency

σ – the electric conductivity

ϵ_r – the dielectric permittivity

The capacitance matrix may be calculated by an electrostatic FEM package.

This time each conductor is defined as a 1V (dc) while the others are set to zero volt. The slot walls are defined as a zero potential boundary. The electrostatic field simulator now solves for the electric potential $\Phi(x,y)$.

$$\nabla(\epsilon_r \epsilon_0 \nabla \Phi(x,y)) = -\rho(x,y) \quad (21.19)$$

$\rho(x,y)$ is the electric charge density.

The result is an $n_c \times n_c$ capacitance matrix per slot which contains the turn-to-turn and turn-to-slot wall capacitance of all conductors in slot. The computation process is done n_c times with always a different single conductor as the 1 V(dc) source.

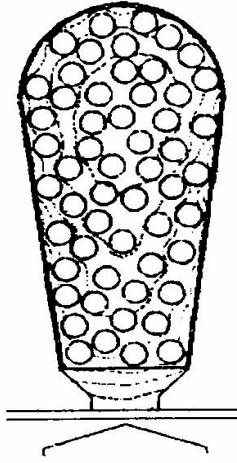


Figure 21.9 Flux distribution in a slot for eddy current FE analysis

The matrix terms C_{ij} are

$$C_{ij} = 2W_{elij} \quad (21.20)$$

W_{elij} is the electric field energy associated to the electric flux lines that connect charges on conductor “i” (active) and j (passive).

As electrostatic analysis is performed the dependence of capacitance on frequency is neglected.

A complete analysis of a motor with 24, 36, 48,...stator slots with each slot represented by $n_c \times n_c$ (for example, 55×55) matrixes would hardly be practical.

A typical line-end coil simulation circuit, with the first 5 individual turns visualised, is shown in [Figure 21.10](#) [7]

To consider extreme possibilities, the line end coil and the last five turns of the first coil are also simulated turn-by-turn. The rest of the turns are simulated by lumped parameters. All the other coils per phase are simulated by lumped parameters. Only the diagonal terms in the impedance and capacitance matrixes are non zero. Saber-simulated voltage drops across the line-end coil and in its first three turns are shown in [Figure 21.11 a](#) and [b](#) for a 750 V voltage pulse with $t_{rise} = 1 \mu s$. [7]

The voltage drop along the line-end coil was 280 V for $t_{rise} = 0.2 \mu s$ and only 80 V for $t_{rise} = 1 \mu s$. Notice that there are 6 coils per phase. Feeder cable tends to lead to 1.2–1.6kV voltage amplitude by wave reflection and thus dangerously high electric stress may occur within the line-end coil of each phase, especially for IGBTs with $t_{rise} < 0.5 \mu s$, and random wound machines.

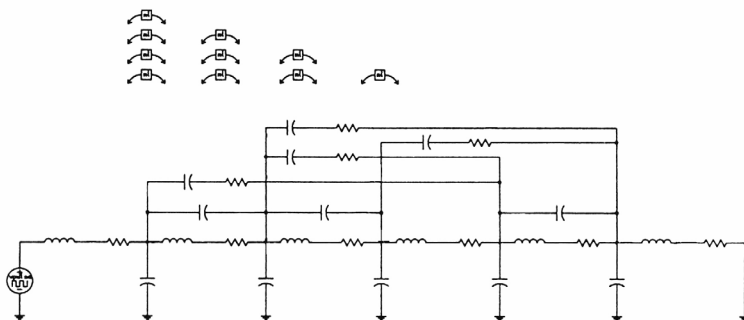
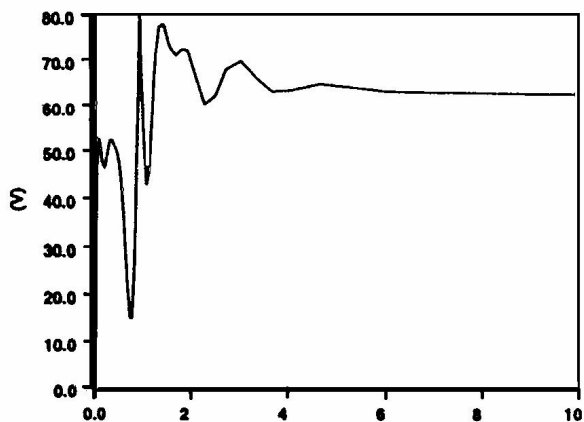


Figure 21.10 Equivalent circuit of the line – end coil

Thorough frequency response measurements with a tapped winding IM have been performed to measure turn–turn and turn to ground distributed parameters. [8]

Further on, the response of windings to PWM input voltages (rise time: $0.24\mu\text{s}$) with short and long cables, have been obtained directly and calculated through the distributed electric circuit with measured parameters. Rather satisfactory but not very good agreement has been obtained. [8]

It was found that the line-end coil (or coil 01) takes up 52% and the next one (coil 02) also takes a good part of input peak voltage (42%). [8] This is in contrast to FE analysis results which tend to predict a lower stress on the second coil [7].



a.)

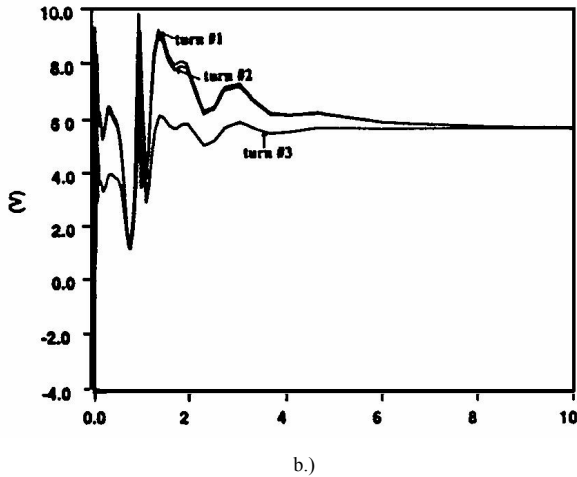


Figure 21.11 Voltage drop versus time for $f_{\text{rise}} = 1 \mu\text{s}$
a) line-end coil; b) Turns 1,2,3 of line- end coil

As expected, long power cables tend to produce higher voltage surges at motor terminals. Consequently, the voltage drop peaks along the line-end coil and its first turn are up to (2–3) times higher. The voltage distribution of PWM voltage surges along the winding coils, especially along its line-end first 2 coils and the line-end first 3 turns thus obtained, is useful to winding insulation design.

Also preformed coils seem more adequate than random coils.

It is recommended that in power-grid fed IMs, the timing between the first, second, and third pole power switch closure be from 0–700 μs . Consequently, even if the commutation voltage surge reaches 5 p.u. [9], in contrast to 2–3 pu for PWM inverters, the voltage drops along the first two coils and their first turns is not necessarily higher because the t_{rise} time of commutation voltage surges is much larger than 0.2–1 μs . As we already mentioned, the second main effect of voltage surges on the IM is the bearing early failures with PWM inverters. Explaining this phenomenon however requires special lumped parameter circuits.

21.6 BEARING CURRENTS CAUSED BY PWM INVERTERS

Rotor eccentricity, homopolar flux effects or electrostatic discharge are known causes of bearing (shaft) ac currents in power grid fed IMs. [10,11] The high frequency common mode large voltage pulse at IM terminals, when fed from PWM inverters, has been suspected to further increase bearing failure. Examination of bearing failures in PWM inverter-fed IM drives indicates fluting, induced by electrical discharged machining (EDM). Fluting is characterized by pits or transverse grooves in the bearing race which lead to premature bearing wear.

When riding the rotor, the lubricant in the bearing behaves as a capacitance. The common mode voltage may charge the shaft to a voltage that exceeds the lubricant's dielectric field rigidity believed to be around $15V_{\text{peak}}/\mu\text{m}$. With an average oil film thickness of 0.2 to 2 μm , a threshold shaft voltage of 3 to 30 V_{peak} is sufficient to trigger electrical discharge machining (EDM).

As already shown in Section 21.1, a PWM inverter produces zero sequence besides positive and negative sequence voltages.

These voltages reach the motor terminals through power cables, online reactors, or common mode chokes. [2] These impedances include common mode components as well.

The behavior of the PWM inverter IM system in the common voltage mode is suggested by the three phase schemata in [Figure 21.12](#). [12]

The common mode voltage, originating from the zero sequence PWM inverter source, is distributed between stator and rotor neutral and ground (frame):

C_{sf} – is the stator winding-frame stray capacitor,

C_{sr} – stator – rotor winding stray capacitor (through airgap mainly)

C_{rf} – rotor winding to motor frame stray capacitor

R_b – bearing resistance

C_b – bearing capacitance

Z_l – nonlinear lubricant impedance which produces intermittent shorting of capacitor C_b through bearing film breakdown or contact point.

With the feeding cable represented by a series/parallel impedance Z_s , Z_p , the common mode voltage equivalent circuit may be extracted from 21.12, as shown in [Figure 21.13](#). R_0 , L_0 are the zero sequence impedances of IM to the inverter voltages. Calculating C_{sf} , C_{sr} , C_{rf} , R_b , C_b , Z_l is still a formidable task.

Consequently, experimental investigation has been performed to somehow segregate the various couplings performed by C_{sf} , C_{sr} , C_{rf} .

The physical construction to the scope implies adding an insulated bearing support sleeve to the stator for both bearings. Also brushes are mounted on the shaft to measure V_{rg} .

Grounding straps are required to short outer bearing races to the frame to simulate normal (uninsulated) bearing operation. ([Figure 21.14](#)) [12]

In region A, the shaft voltage V_{rg} charges to about $20V_{\text{pk}}$. At the end of region A, V_{sng} jumps to a higher level causing a pulse in V_{rg} . In that moment, the oil film breaks down at $35 V_{\text{pk}}$ and a $3 A_{\text{pk}}$ bearing current pulse is produced. At high temperatures, when oil film thickness is further reduced, the breakdown voltage (V_{rg}) pulse may be as low as 6–10 volts.

Region B is without bearing current. Here, the bearing is charged and discharged without current.

Region C shows the rotor and bearing (V_{rg} and V_{sng}) charging to a lower voltage level. No EDM occurs this time. $V_{\text{rg}} = 0$ with V_{sng} high means that contact asperities are shorting C_b .

The shaft voltage V_{rg} , measured between the rotor brush and the ground, is a strong indicator of EDM potentiality. Test results on [Figure 21.15](#) [2] show V_{rg} , the bearing current I_b and the stator neutral to ground voltage V_{sng} .

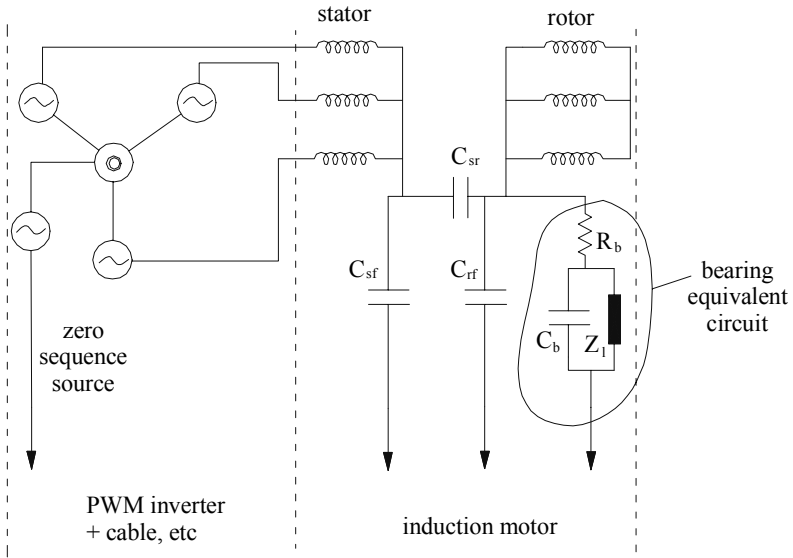


Figure 21.12 Three phase PWM inverter plus IM model for bearing currents

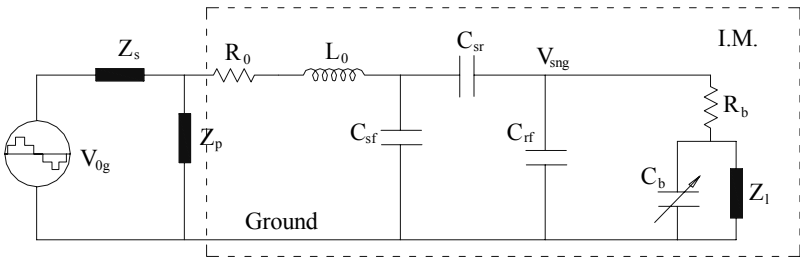


Figure 21.13 Common mode lumped equivalent circuit

An indicator of shaft voltage is the bearing voltage ratio (BVR):

$$\text{BVR} = \frac{V_{\text{rg}}}{V_{\text{sng}}} = \frac{C_{\text{sr}}}{C_{\text{sr}} + C_{\text{b}} + C_{\text{rf}}} \quad (21.21)$$

With insulated bearings, neglecting the bearing current (if the rotor brush circuits are open and the ground of the motor is connected to the inverter frame), the ground current I_{G} refers to stator winding to stator frame capacitance C_{sf} (Figure 21.16 a). With an insulated bearing, but with both rotor brushes connected to the inverter frame, the measured current I_{AB} is related to stator winding to rotor coupling (C_{sr}). (Figure 21.16 b)

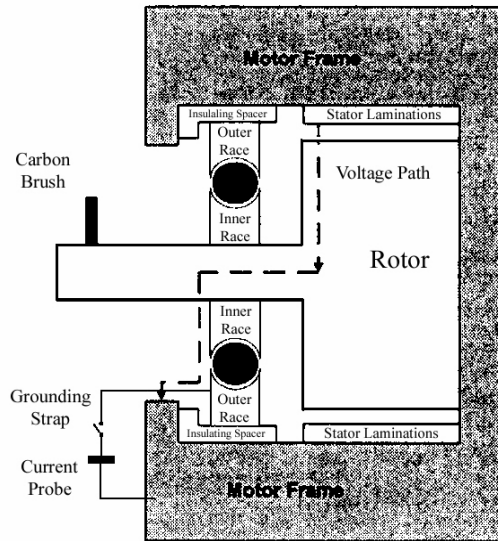


Figure 21.14 The test motor

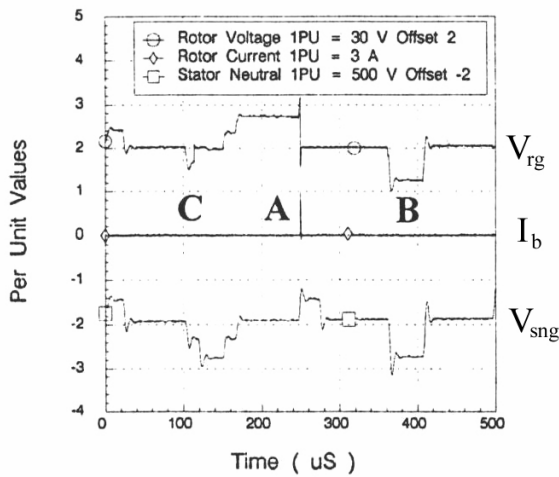


Figure 21.15 Bearing breakdown parameters

In contrast, shortcircuiting the bearing insulation sleeve allows the measurement of initial (uninsulated) bearing current I_b (Figure 21.16 c).

Experiments as those suggested in Figure 21.16 [13] may eventually lead to C_{sr} , C_{sf} , C_{rf} , R_b , C_b identification.

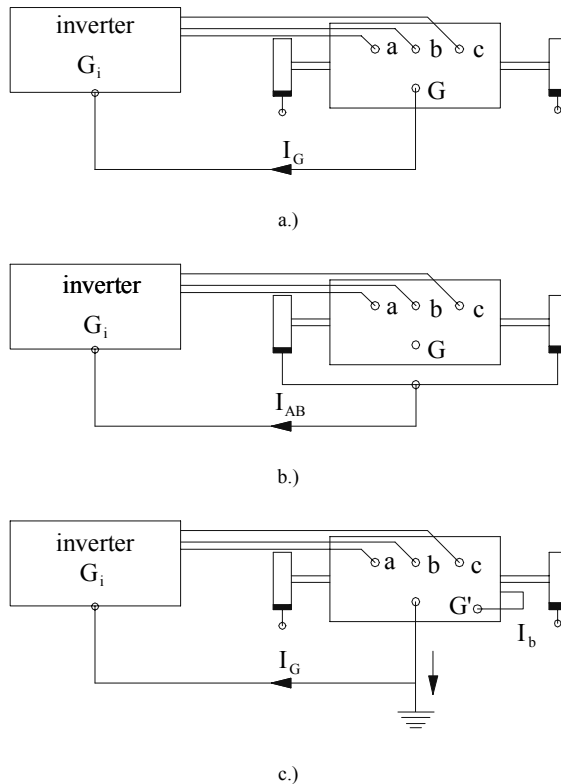


Figure 21.16 Capacitive coupling modes

- a.) stator winding to stator
- b.) stator winding to rotor
- c.) uninsulated bearing current I_b

21.7 WAYS TO REDUCE PWM INVERTER BEARING CURRENTS

Reducing the shaft voltage V_{rg} to less than $1-1.5 V_{pk}$ is apparently enough to avoid EDM and thus eliminate bearing premature failure.

To do so, bearing currents should be reduced or their path be bypassed by larger capacitance path; that is, increasing C_{sf} or decreasing C_{sr} .

Three main practical procedures have evolved [12] so far

- Properly insulated bearings (C_b decrease)
- Conducting tape on the stator in the airgap (to reduce C_{sr})
- Copper slot stick covers (or paint) and end windings shielded with nomex rings and covered with copper tape and all connected to ground (to reduce C_{sr}) – [Figure 21.17 b](#)

Various degrees of shaft voltage attenuation rates (from 50% to 100%) have been achieved with such methods depending on the relative area of the shields. Shaft voltages close to NEMA specifications have been obtained.

The conductive shields do not notably affect the machine temperatures.

Besides the EDM discharge bearing current, a kind of circulating bearing current that flows axially through the stator frame has been identified. [14] Essentially a net high frequency axial flux is produced by the difference in stator coil end currents due to capacitance current leaks along the stack length between conductors in slots and the magnetic core. However, the relative value of this circulating current component proves to be small in comparison with EDM discharge current.

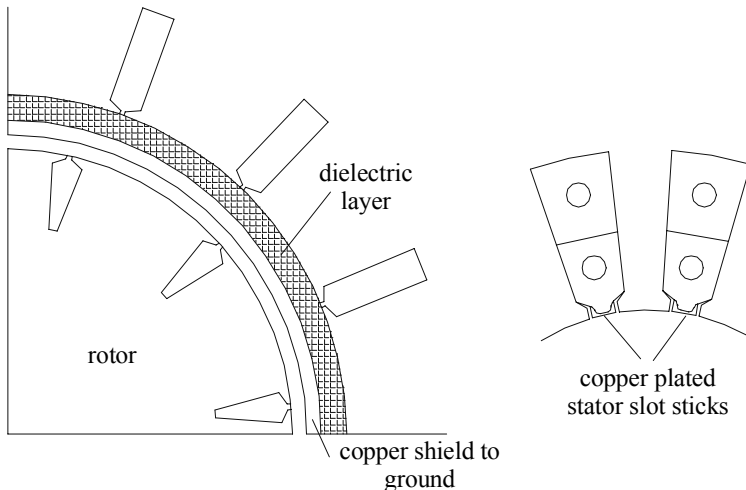


Figure 21.17 Conductive shields to bypass bearing currents (C_{sr} is reduced)

21.8 SUMMARY

- Super-high frequency models for IMs are required to assess the voltage surge effects due to switching operations or PWM inverter-fed operation modes.
- The PWM inverter produces 0.2–2 μ s rise time voltage pulses in both differential of common modes. These two modes seem independent of each other and, due to multiple reflection with long feeder cables, may reach up to 3 p.u. dc voltage levels.
- The distribution of voltage surges within the IM-repetitive in case of PWM inverters – along the stator windings is not uniform. Most of the voltage drops along the first two line-end coils and particularly along their first 1-3 turns.

- The common mode voltage pulses seem to produce also premature bearing failure through electric discharge machining (EDM). They also produce most of the electromagnetic interference effects.
- To describe the global response of IM to voltage surges, the differential impedance Z_m , the neutral impedance Z_{ng} , and the common voltage (or ground) impedance Z_{og} are defined through pertinent stator winding connections (Figure 21.3) in order to be easily estimated through direct measurements.
- The frequency response tests performed with RLC analyzers suggest simplified lumped parameter equivalent circuits.
- For the differential impedance Z_m , a resonant parallel circuit with high frequency capacitance C_{hf} and eddy current resistance R_{hf} in series is produced. The second R_{lf} , L_{lf} branch refers to lower frequency modelling (1 kHz to 50 kHz or so). Such parameters are shown to vary almost linearly with motor power (Table 21.1). Care must be exercised, however, as there may be notable differences between $Z_m(p)$ from different manufacturers, for given power.
- The neutral (Z_{ng}) and ground (common voltage) Z_{og} impedances stem from the phase lumped equivalent circuit for high frequencies (Figure 21.8). Its components are derived from frequency responses through some approximations.
- The phase circuit boils down to two ground capacitors (C_g) with a high frequency inductance in between, L_d . C_d and L_d are shown to vary simply with motor power ((21.15)-(21.16))
- To model the voltage surge distribution along the stator windings, a distributed high frequency equivalent circuit is necessary. Such a circuit should visualize turn-to-turn and turn-to-ground capacitances, turn-to-turn inductances and eddy current resistances. The computation of such distributed parameters has been attempted by FEM, for super-high frequencies (1 MHz or so). The electromagnetic and electrostatic field is non zero only in the stator slots and in and around stator coil end connections.
- FEM-derived distributed equivalent circuits have been used to predict voltage surge distribution for voltage surges within 0-2 μs rising time. The first coil takes most of the voltage surge. Its first 1-3 turns particularly so. The percentage of voltage drop on the line-end coil decreases from 70% for 0.1 μs rising time voltage pulses to 30% for 1 μs rising time. [15]
- Detailed tests with thoroughly tapped windings have shown that as the rising time t_{rise} increases not only the first but also the second coil takes a sizeable portion of voltage surge. [8] Improved discharge resistance (by adding oxides) insulation of magnetic wires [1], together with voltage surge reduction are the main avenues to long insulation life in IMs.

- Bearing currents, due to common mode voltage surges, are deemed responsible for occasional bearing failures in PWM inverter-fed IMs, especially when fed through long power cables.
- The so-called shaft voltage V_{rg} is a good indicator of bearing-failure propensity. In general, V_{rg} values of 15 V_{pk} at room temperature and 6–10 V_{pk} in hot motors are enough to trigger the electrostatic discharge machining (EDM) which produces bearing fluting. In essence, the high voltage pulses reduce the lubricant non linear impedance and thus a large bearing current pulse occurs which further advances the fluting process towards bearing failure.
- Insulated bearings or Faraday shields (in the airgap or on slot taps, covering also the end connections) are practical solutions to avoid large bearing EDM currents. They increase the bearings life.
- Given the complexity and practical importance of super-high frequency behavior of IMs, much progress is expected in the near future, including pertinent international test standards.

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